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TITLE – A Method for Performing a Bit Error Rate Analysis of the ALSEP Telemetry Data Link.

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ABSTRACT

A mathematical model of the Apollo MSFN telemetry receiver used to detect ALSEP telemetry data is defined. This model is analyzed to determine power margins for a bit error probability $P_E = 10^{-4}$ for each possible combination of system parameters that may be used. In particular, results are obtained which allow power margins to be determined for telemetry bit rates of 530, 1060 and 10,600 bps when the carrier channel predetection bandwidth and the two-sided threshold carrier tracking loop bandwidth combinations are 2KHz and 12 Hz and 7 KHz and 50 Hz. The receiver model used accounts for a non-perfect carrier phase reference used in the demodulation process that is derived in the phase locked loop (carrier channel) in the ground system. However, it is found that a perfect carrier reference can be assumed when the receiver predetection bandwidth is 2 KHz and the phase locked loop two-sided threshold bandwidth is 12 Hz. To demonstrate the method developed in the memorandum, the performance margins for the ALSEP telemetry link are determined under nominal operating conditions.

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TECHNICAL MEMORANDUM

1.0 INTRODUCTION

The purpose of this memorandum is to describe and analyze a mathematical model of the telemetry receiver at MSFN sites used to demodulate data from the Apollo Lunar Surface Experiment Package (ALSEP). In this analysis it is desired to compute power margins that exist when the probability of error in detecting a data bit is 10^{-4}. There are a variety of possible combinations of bit rates and circuit parameters that may exist in the telemetry channel. For each combination of these parameters, power margins must be determined for the desired maximum error rate of 10^{-4}. A phase locked loop is used in the receiver to obtain a carrier reference which is used for coherent product detection of the split-phase PCM telemetry data train. When the carrier phase is assumed to be exactly determined, the calculation of margins is simple. However, if the carrier reference is not perfect the margin calculations are much more involved. In the work discussed here the effect of a nonperfect carrier reference is included. In some possible modes of operation the effect of the carrier reference phase error is so small that a perfect reference can be assumed. These situations are discussed.

2.0 DISCUSSION

2.1 Limiter-Phase Locked Loop Configuration Used to Acquire Carrier Reference

To acquire the carrier reference for coherent data demodulation, a phase locked loop (PLL) is used. The PLL is preceded by a bandpass limiter that removes AM from the input PM signal. The bandpass limiter is, in turn, preceded by a bandpass filter denoted as the prefilter in Figure 1.

Both the postfilter, used in the bandpass limiter, and the prefilter have center frequency w_c and bandwidth B_c. The gains of the two filters are normalized to unity in their pass bands. The bandpass limiter is defined as a hard limiter followed by a bandpass filter. The hard limiter used here is normalized to ± 1 limits.

2.2 Limiter Characteristics

The output-to-input S/N relationship and the signal voltage suppression factor α for an ideal symmetrical bandpass limiter are well known[(1)(2)]. When $S/N)_{in} \to 0$, then $(S/N)_{out} \cong \frac{\pi}{4}(S/N)_{in}$, and when $(S/N)_{in} \to \infty$, $(S/N)_{out} = 2(S/N)_{in}$. However, care must be taken when applying these results to a bandpass limiter in cascade with a PLL. There is a 3db improvement in the limiter when $(S/N)_{in}$ is large. There is also a 3db S/N improvement in a PLL. The improvement in the limiter is due to the removal of the in-phase noise component of the additive noise present at the limiter input. The 3db improvement in the PLL is also due to removal of the in-phase noise component present at the PLL input. Since the individual improvements are due to the same source a 6db improvement in (S/N) ratio can not be expected when the limiter and PLL are used in cascade.

For the limiter - PLL cascade used here the limiter asymptotes are assumed to be

$$\left(\frac{S}{N}\right)_{out} = \frac{\pi}{4}\left(\frac{S}{N}\right)_{in} \quad \text{for} \quad \left(\frac{S}{N}\right)_{in} \to 0$$

and

$$\left(\frac{S}{N}\right)_{out} \cong \left(\frac{S}{N}\right)_{in} \quad \text{for} \quad \left(\frac{S}{N}\right)_{in} \to \infty$$

The (S/N) transfer characteristic for the bandpass limiter when in cascade with a PLL has been approximated by the piecewise linear equation

$$\left(\frac{S}{N}\right)_{out} = \frac{\pi}{4}\left(\frac{S}{N}\right)_{in} + (0.682)u_{-1}\left\{\left(\frac{S}{N}\right)_{in} - 0.035\right\}\cdot\left[\left(\frac{S}{N}\right)_{in} - 0.035\right]\cdot\left(\frac{S}{N}\right)_{in}$$

$$- (0.682)u_{-1}\left\{\left(\frac{S}{N}\right)_{in} - 0.35\right\}\cdot\left[\left(\frac{S}{N}\right)_{in} - 0.35\right]\cdot\left(\frac{S}{N}\right)_{in} = F\left[(S/N)_{in}\right]$$

where

$$u_{-1}\{x-a\} = \begin{cases} 1, & x \geq a \\ 0, & x < a \end{cases}$$

This functional relationship is shown in Figure 2 in terms of $(S/N)_{out}/(S/N)_{in}$ vs $(S/N)_{in}$.

The $(S/N)_{in}$ in terms of $(S/N)_{out}$ follows directly from $(S/N)_{out} = F[(S/N)_{in}]$. It follows that if

$$\left(\frac{S}{N}\right)_{in} \cong \frac{4}{\pi}\left(\frac{S}{N}\right)_{out} \qquad \text{for} \qquad \left(\frac{S}{N}\right)_{in} \leq 0.035$$

then this approximation also holds if

$$(S/N)_{out} \leq \left(\frac{\pi}{4}\right)(0.035) = 0.0275$$

and if

$$\left(\frac{S}{N}\right)_{in} \cong \left(\frac{S}{N}\right)_{out} \qquad \text{for} \qquad (S/N)_{in} \geq 0.35$$

then this approximation also will hold for

$$(S/N)_{out} \geq 0.35$$

Therefore,

$$(S/N)_{in} = \frac{4}{\pi}\left(\frac{S}{N}\right)_{out} - (0.837)u_{-1}\left\{\left(\frac{S}{N}\right)_o - 0.0275\right\}\cdot\left[\left(\frac{S}{N}\right)_o - 0.0275\right]\cdot\left(\frac{S}{N}\right)_o$$

$$+ (0.837)u_{-1}\left\{\left(\frac{S}{N}\right)_o - 0.35\right\}\cdot\left[\left(\frac{S}{N}\right)_o - 0.35\right]\cdot\left(\frac{S}{N}\right)_o = f\left[\left(\frac{S}{N}\right)_o\right]$$

The piecewise linear graph for $\frac{(S/N)_{in}}{(S/N)_{out}}$ vs $\left(\frac{S}{N}\right)_{out}$ is shown in Figure 3.

The signal attenuation of the bandpass limiter is

$$\alpha = \left(\frac{\text{Signal Component Out}}{\text{Total Limiter Input Power}}\right)$$

$$\cong \frac{1}{\left[1 + \frac{4}{\pi}\left(\frac{N}{S}\right)_{in}\right]^{1/2}}$$

where $(S/N)_{in}$ is defined in Figure 1 as the limiter input signal-to-noise power ratio.

2.3 Carrier Signal-to-Noise Power Ratio In The $2B_L$ Loop Bandwith

For the carrier modulated by a split phase binary waveform, the total power in the IF signal $(S)_{IF}$ and the power in the carrier component of this signal, $(S)_{IF_C}$ are related by

$$(\cos^2 \theta)\ (S)_{IF} = (S)_{IF_C}$$

where θ is the phase modulation index. For a given value of B_C the amount of signal power removed from the limiter input by the prefilter is determined by the bit rate H. For a given bit rate and bandwidth B_C the total signal power into the limiter $(S)_{in}$ is related to the IF signal power, $(S)_{IF}$, by

$$k(S)_{IF} = (S)_{in} \text{ where } 0 \leq k \leq 1$$

The carrier component of the IF signal is not attenuated by the prefilter. Hence

$$(S)_{in_C} = (S)_{IF_C} \quad .$$

Then

$$(S)_{In_C} = \left(\frac{\cos^2 \theta}{k}\right) (S)_{In}$$

The split-phase modulating waveform is such that its baseband spectrum is

$$G[f] = \left[\frac{V^2}{H} \frac{\sin^4 (\omega/4H)}{(\omega/4H)^2}\right]$$

where H is the bit rate. The spectrum is shown in Figure 4.

The low pass characteristic of the PLL is shown on the same figure as G(f). Because of the relatively narrow bandwidth, B_L, of the PLL, only the carrier component of the signal received at the PLL effects the loop operation when the loop is in lock. Then

$$\left(\frac{S}{N}\right)_{2B_L} = \left(\frac{\cos^2 \theta}{k}\right) \left(\frac{B_C}{2B_L}\right) \left(\frac{S}{N}\right)_{out}$$

Since

$$\left(\frac{S}{N}\right)_{in} = f\left[\left(\frac{S}{N}\right)_{out}\right]$$

as discussed above,

$$\left(\frac{S}{N}\right)_{in} = f\left[\frac{k}{\cos^2 \theta} \cdot \frac{2B_L}{B_C} \left(\frac{S}{N}\right)_{2B_L}\right]$$

and

$$\alpha = \left[1 + \frac{4}{\pi}\left(\frac{N}{S}\right)_{in}\right]^{-1/2}$$

$$= \left[1 + \frac{\left(\frac{4}{\pi}\right)}{f\left[\left(\frac{k}{\cos^2\theta}\right)\left(\frac{2B_{L0}}{B_C}\right)\left(\frac{S}{N}\right)_{2B_{L0}}\right]}\right]^{-1/2}$$

Loop conditions in $2B_{L0}$ may be substituted here, since

$$\left(\frac{N}{S}\right)_{2B_L} = \left(\frac{B_L}{B_{L0}}\right)\left(\frac{N}{S}\right)_{2B_{L0}}$$

It is assumed that the loop bandwidth B_L is given by

$$B_L = \frac{B_{L0}}{3}\left(1 + \frac{2\alpha}{\alpha_o}\right)$$

when the entire IF signal spectrum falls into the passband of the PLL. Here

$$\alpha_o = \left[1 + \frac{4}{\pi}\left(\frac{N}{S}\right)_{in}\right]^{-1/2}$$

when $\left(\frac{N}{S}\right)_{in}$ corresponds to $(N/S)_{2B_{L0}} = 1$. Since the carrier component of the signal into the loop is all that effects the loop bandwidth and since this signal voltage is $\cos\theta$ times the loop input signal, a new B_L expression is obtained from the original by replacing α with $\alpha \cos\theta$. Then for the ALSEP telemetry receiver in the MSFN,

$$B_L = \frac{B_{L0}}{3}\left(1 + \frac{2\alpha\cos\theta}{\alpha_o}\right)$$

The signal-to-noise ratio (γ) in the passband of the PLL is

$$\gamma = \frac{P_C}{B_L N_o}$$

where

P_C = carrier power into loop,

B_L = noise equivalent bandwidth of the PLL (one-sided) and

N_o = one-sided noise spectral density at the loop input.

The signal-to-noise ratio in the $2B_{L0}$ bandwidth is $\frac{P_C}{2B_{L0}N_o}$. Then in terms of $(S/N)_{2B_{L0}}$, γ is given by

$$\gamma = \frac{6}{\left[1 + \frac{2\alpha\cos\theta}{\alpha_o}\right]} \cdot \left(\frac{S}{N}\right)_{2B_{L0}}$$

2.4 The Effect of Non-Perfect Carrier Reference on Demodulation

The term γ is used by W. C. Lindsey to determine the bit error probability P_E vs (S/N) in the bit rate bandwidth, $(S/N)_{BRB}$, when a non-perfect carrier reference is obtained within the PLL[3]. For a given signal-to-noise ratio in $2B_L$, a lower limit for P_E is reached as $\left(\frac{S}{N}\right)_{BRB}$ increases without bound. For a perfect reference, this is not the case. If $\gamma = \infty$, P_E decreases indefinitely with $(S/N)_{BRB}$. The conclusions of (3) are summarized in Figure 5 for the one-way communication link. Bit error, P_E, is given as a function of $(S/N)_{BRB}$ with signal-to-noise in $2B_L$ loop bandwidth as a parameter. Using Figure 5, a family of constant probability curves has been determined. For fixed values of P_E, $(S/N)_{BRB}$ is plotted as a function of γ_{db} as shown in Figure 6.

2.5 The Relationship of $(S/N)_{2B_{LO}}$ and $(S/N)_{BRB}$

Thus far two relationships have been determined. (1) γ_{db} vs $(S/N)_{2B_{LO}(db)}$ and (2) $(S/N)_{BRB(db)}$ vs γ_{db} for given P_E. A third relationship between $(S/N)_{BRB(db)}$ and $(S/N)_{2B_{LO}(db)}$ exist independently of the others. In the carrier channel the limiter input signal-to-noise ratio is

$$\left(\frac{S}{N}\right)_{in} = \frac{k\ P_r}{B_C\ N_o}$$

where P_r is the IF signal power at the prefilter and N_o is the noise density (one-sided) at the IF. From before,

$$\left(\frac{S}{N}\right)_{2B_{LO}} = \left(\frac{S}{N}\right)_{out} \cdot \left(\frac{B_C}{2B_{LO}}\right)\left(\frac{\cos^2\theta}{k}\right)$$

since the signal in $2B_{LO}$ is the carrier component. The limiter effect on the S/N ratio is described by the equivalent equations

$$\left(\frac{S}{N}\right)_{out} = F\left[\left(\frac{S}{N}\right)_{in}\right] \quad \text{and} \quad \left(\frac{S}{N}\right)_{in} = f\left[\left(\frac{S}{N}\right)_{out}\right]$$

where F and f were defined above for the limiter - PLL cascade. Then

$$\left(\frac{S}{N}\right)_{2B_{LO}} = F\left[\frac{k\ P_r}{B_C\ N_o}\right] \cdot \frac{B_C}{2B_{LO}} \cdot \frac{\cos^2\theta}{k}$$

For the data channel, the S/N ratio at limiter input is

$$\left(\frac{S}{N}\right)_{in_D} = \frac{P_r}{N_o B_D}$$

But

$$\left(\frac{S}{N}\right)_{BRB} = \left(\frac{S}{N}\right)_{out_D} \cdot \frac{B_D}{B_H} \cdot \sin^2\theta \cdot L_{DE}$$

where L_{DE} is the decommutator loss factor. Since

$$\left(\frac{S}{N}\right)_{out_D} = F\left[\left(\frac{S}{N}\right)_{in_D}\right] \quad ,$$

it follows that

$$\left(\frac{S}{N}\right)_{BRB} = F\left(\frac{P_r}{N_o B_D}\right) \cdot \frac{B_D}{B_H} \cdot \sin^2\theta \cdot L_{DE}$$

and

$$F\left(\frac{P_r}{N_o B_D}\right) = \left(\frac{S}{N}\right)_{BRB} \cdot \left(\frac{B_D}{B_H} \sin^2\theta \; L_{DE}\right)^{-1}$$

From before, F and f are such that for any x and y related by

$$x = F(y)$$

it must follow that

$$y = f(x)$$

Then

$$\frac{P_r}{N_o} = B_D \cdot f\left\{\left(\frac{S}{N}\right)_{BRB} \cdot \left(\frac{B_D}{B_H} \sin^2\theta \; L_{DE}\right)^{-1}\right\}$$

Therefore, $(S/N)_{2B_{LO}}$ and $(S/N)_{BRB}$ are related by

$$\left(\frac{S}{N}\right)_{2B_{LO}} = F\left[\frac{k}{B_C} \cdot B_D \; f\left\{\left(\frac{S}{N}\right)_{BRB} \cdot \left(\frac{B_D}{B_H} \sin^2\theta \; L_{DE}\right)^{-1}\right\}\right] \cdot \frac{B_C}{2B_{LO}} \cdot \frac{\cos^2\theta}{k}$$

If $f(x) = ax$ and $F(y) = by$ where a and b are constants the relationship reduced to

$$\left(\frac{S}{N}\right)_{2B_{LO}} = \left(\frac{S}{N}\right)_{BRB} \left(b \; \frac{k}{B_C} \; B_D \; a\right) \left(\frac{B_D}{B_H} \sin^2\theta \; L_{DE}\right)^{-1} \left(\frac{B_C}{2B_{LO}} \; \frac{\cos^2\theta}{k}\right)$$

Since $x = F(y)$ and $y = f(x)$, a and b must be related such that $x = by$ and $y = ax$. That is, $a = 1/b$, if the two limiters are identical. Then

$$\left(\frac{S}{N}\right)_{2B_{LO}(db)} = \left(\frac{S}{N}\right)_{BRB(db)} + \left(\frac{B_H}{2B_{LO}}\right)_{db} - (\tan^2\theta)_{db} - L_{DE_{db}}$$

In the equations derived thus far for γ vs $(S/N)_{2B_{LO}}$ and $\left(\frac{S}{N}\right)_{2B_{LO}}$ vs $\left(\frac{S}{N}\right)_{BRB}$, approximating assumptions have been avoided. The equations are difficult to handle when the variation of limiter $(S/N)_{out}$ vs $(S/N)_{in}$ is incorporated. In the foregoing analysis the variation (approximately 1db) in the limiter (S/N) transfer characteristic is not considered. It will be assumed that $(S/N)_{in} = (S/N)_{out}$ for all $(S/N)_{in}$ values both for the carrier channel limiter and the data channel limiter. It is considered that the errors introduced by this approximation are small enough to be neglected.

2.6 Calculation of Signal to Noise Ratio (γ) in Passband of PLL vs. $(S/N)_{2B_{LO}}$

For ALSEP, plots of γ_{db} vs $(S/N)_{2B_{LO}}$ were made for two sets of system parameters. They are

(I)	$\theta = 1.25$ rad.	(II)	$\theta = 1.25$ rad.
	$\alpha_o = 0.0685$		$\alpha_o = 0.0708$
	$B_C = 2$ KHz		$B_C = 7$ KHz
	$2B_{LO} = 12$ Hz		$2B_{LO} = 50$ Hz

In each case the data bandwidth B_D is assumed to be sufficiently large so as to pass the signal undistorted. In each case two values of bit rate H are considered. These are H = 530 bps and H = 1060 bps. When H = 1060 in case (I), the prefilter of $B_C = 2$ KHz removes 30% of the IF signal power before the limiter. Hence when H = 1060 in (I), k = 0.7. However, for H = 530 in (I) and H = 530, 1060 in (II), k = 1.

In case (I) the results of calculation are with

k = 1.0 (H = 530 bps)

$\left(\frac{S}{N}\right)_{2B_{LO}(db)}$	γ_{db}
0	3.06
10	9.82
20	18.03
30	27.67
40	37.65

with

$k = 0.7$ $(H = 1060$ bps$)$	
$\left(\frac{S}{N}\right)_{2B_{L0}(db)}$	γ_{db}
0	3.54
10	10.28
20	18.18
30	27.72
40	37.67

In case (II) the results of calculation are with

$k = 1$ for $H = 530, 1060$	
$\left(\frac{S}{N}\right)_{2B_{L0}(db)}$	γ_{db}
0	2.92
10	9.75
20	18.14
30	27.84
40	37.81

Figure 7 gives γ_{db} vs $(S/N)_{2B_{L0}}$ for $k = 1, 0.7$ in case (I). Figure 8 gives γ_{db} vs $(S/N)_{2B_{L0}}$ for $k = 1$ in both cases (I) and (II). In both cases I and II when $k = 1$ as well as case I when $k = 0.7$, the graphs are closely represented by the linear equation

$$(0.83)(S/N)_{2B_{L0}(db)} + 1.55 = \gamma_{db}$$

for values of γ_{db} between 5 db and 20 db. This can be combined with the equation of $(S/N)_{BRB}$ vs $(S/N)_{2B_{L0}}$. With the approximation $(S/N)_{in} = (S/N)_{out}$ for the limiters, the resulting equation obtained by elimination of $(S/N)_{2B_{L0}}$ is

$$\left(\frac{S}{N}\right)_{BRB(db)} = 1.205\ \gamma_{db} - 1.87 - \left(\frac{B_H}{2B_{LO}}\right)_{db}$$

$$+ (\tan^2\theta)_{db} + L_{DE_{db}}$$

For given bit rate bandwidth, B_H, $2B_{LO}$, θ and L_{DE}, this linear equation in db can be plotted directly on the family of constant P_E curves in Figure 6. For H = 530 bps, θ = 1.25 rad., $2B_{LO}$ = 12 Hz and $L_{DE_{(db)}}$ = -1 db, the equation is

$$\left(\frac{S}{N}\right)_{BRB(db)} = 1.205\ \gamma_{db} - 9.78$$

For H = 1060 bps, θ = 1.25, $2B_{LO}$ = 12 Hz and $L_{DE_{db}}$ = -1,

$$\left(\frac{S}{N}\right)_{BRB(db)} = 1.205\ \gamma_{db} - 12.78$$

For H = 530 bps, θ = 1.25, $2B_{LO}$ = 50 Hz and $L_{DE_{db}}$ = -1,

$$\left(\frac{S}{N}\right)_{BRB(db)} = 1.205\ \gamma_{db} - 3.58$$

For H = 1060 bps, θ = 1.25, $2B_{LO}$ = 50 Hz and $L_{DE_{db}}$ = -1,

$$\left(\frac{S}{N}\right)_{BRB(db)} = 1.205\ \gamma_{db} - 6.58$$

When $L_{DE_{db}}$ = -3 the corresponding equations are obtained by subtracting -2 db from the right side of each of the above linear equations. The graphs of the four equations of $(S/N)_{BRN(db)}$ vs γ_{db} for L_{DE} = -1 db are shown on the constant P_E family of Figure 6.

The equations of $(S/N)_{BRB(db)}$ vs γ_{db} plotted in Figure 6 are equations of restraint imposed by the system, whereas the constant P_E curves relate γ_{db} and $(S/N)_{BRB(db)}$ for specific values of P_E. For a given probability of bit error, P_E, the values of $(S/N)_{BRB}$ is fixed for given H, B_{LO}, θ and L_{DE}. Since $\frac{P_r}{N_o}$ is directly related to $(S/N)_{BRB}$ in a one-to-one fashion, $\frac{P_r}{N_o}$ is also fixed for given P_E and given parameter values. The ratio P_r/N_o is found from

$$\left(\frac{P_r}{N_o}\right)_{db} = \left(\frac{S}{N}\right)_{BRB(db)} + (B_H)_{db} - (\sin^2\theta)_{db} - L_{DE_{db}}$$

under the approximation of unity transfer for S/N ratio in the limiters.

2.7 Communication Performance Margins For The ALSEP Telemetry Link

To demonstrate the utility of the method developed here, the performance margins for the ALSEP Telemetry Link have been computed assuming a 30 foot MSFN antenna is used.

2.7.1 Calculation of Received Signal Power (See Reference 4)

	Parameter	Nominal Value	Tolerances +db	-db
(1)	ALSEP Transmitter Power (dbm)	+ 30.0	1	0
(2)	ALSEP Transmitter Ckt. Loss (db)	- 2.4	0	0.3
(3)	ALSEP Antenna Gain (db)	+ 15.8	0.5	0.6
(4)	ALSEP Antenna Pointing Loss (db)	- 3.0	3.0	0.7
(5)	Space Loss (db)	-211.6	0.4	0.2
(6)	Polarization Loss (db)	- 0.2	0.2	0.3
(7)	Multipath Loss (db)	0	0	1.0
(8)	MSFN Antenna Gain (30' Dish)(db)	+ 44.0	0	0
(9)	MSFN Antenna Pointing Loss (db)	0	0	0
(10)	MSFN Receiver Ckt. Loss (db)	Included in Antenna Gain		
(11)	MSFN Receiver Input Power for 30' Dish (dbm),(Sum of 1 - 10)	-127.4	5.1	3.1

2.7.2 Calculation of Receiver Noise Spectral Density (see Reference 4, page 20)

	For the 30' Antenna:	N_o
(1)	Moon at zenith, cooled paramp (dbm/Hz)	-176.1
(2)	Moon at zenith, uncooled paramp (dbm/Hz)	-174.6
(3)	Moon at horizon uncooled paramp (dbm/Hz)	-173.1

The signal-to-noise power ratio in the bit rate bandwidth $(S/N)_{BRB}$ is related to the receiver input signal power P_r and input noise spectral density N_o by the equation

$$\left(\frac{P_r}{N_o}\right)_{db} = \left(\frac{S}{N}\right)_{BRB(db)} + (B_H)_{db} - (\sin^2\theta)_{db} - L_{DE(db)}$$

For the nominal values of P_r =-127.4 dbm and N_o = -174.6 dbm;

$$\left(\frac{P_r}{N_o}\right)_{db} = 47.2 \text{ db}$$

with θ = 1.25 radians, $(\sin^2\theta)_{db}$ = -0.46. With a decommutator loss of -1 db, $- L_{DE}$ = + 1 db. Then

$$\left(\frac{S}{N}\right)_{BRB(db)} = 47.2 + (-0.46) + (-1) - (B_H)_{db} = 45.74 - (B_H)_{db}$$

For a data bit rate of 1060 bps, B_H = 1060 and $(B_H)_{db} = 10 \log_{10} (1.06 \times 10^3) = 30 + 10 \log_{10} 1.06 = 30.26$. Then under nominal operating conditions

$$\left.\frac{S}{N}\right)_{BRB(db)} = 15.48 \text{ db}$$

The margin in $(S/N)_{BRB(db)}$ for $P_E = 10^{-4}$ will be obtained for the two bandwidth combinations B_C = 7 KHz with $2B_{L0}$ = 50 Hz, and B_C = 2 KHz with $2B_{L0}$ = 12 Hz. With reference to Figure 6, the intersection of the $P_E = 10^{-4}$ curve with the restraint curve for B_C = 7 KHz and $2B_{L0}$ = 50 Hz occurs at $(S/N)_{BRB(db)}$ = 8.7 db. The intersection of the $P_E = 10^{-4}$ curve and the restraint curve for $B_C = 2$ KHz and $2B_{LO} = 12$ Hz occurs when $(S/N)_{BRB(db)} = 8.4$ db.

With the data rate H = 1060 bps and $B_C = 7$ KHz with $2B_{LO} = 50$ Hz the margin in $(S/N)_{BRB(db)}$ is 15.48 - 8.7 = 6.78 db. With the data rate of H = 1060 bps and $B_C = 2$ KHz with $2B_{LO} = 12$ Hz, the margin is slightly greater at 7.08 db.

The difference between the margins calculated under the assumption of a perfect carrier reference and those obtained when the non-perfect reference is taken into account is slight for the system parameters considered. This can be seen from Figure 6.

3.0 CONCLUSIONS

The constant probability of error (P_E) curves presented in Figure 6 approach the classic definition of bit error rate when there is a high signal-to-noise ratio (γ) in the carrier tracking phase locked loop. For ALSEP, when nominal system parameters are assumed (i.e., modulation indices, loop bandwidths etc.) it is found that the performance of the System can be predicted by considering the carrier phase to be exact. If ground station receiver bandwidths, similar to those used with the Apollo S-band system, are employed, particularly for the low ALSEP data rate (i.e., 530 BPS), it is necessary to consider the imperfect carrier reference.

Curves are presented in Figure 6 that enable the determination of (1) the required signal power to noise spectral $\left(\frac{P_r}{N_o}\right)$ for a given bit error rate, and (2) if $\left(\frac{P_r}{N_o}\right)$ is known, the margin for both the carrier and data channels. The margins for the carrier and data channels for a specified bit error rate and System configuration are shown to be related (i.e., not independent).

In general, it follows that the larger the ratio of the bit rate bandwidth to $2B_{LO}$, the more valid the perfect reference approximation becomes.

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2021-WDW-jdc

Attachments
References
Figures 1 thru 8

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REFERENCES

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3. Lindsey, W. C., "The Effects of Radio Frequency (RF) Timing Noise in Two-Way Communication Systems", Space Programs Summary No. 37-32, Volume IV, Jet Propulsion Laboratory, Pasadena, California, April 30, 1965.

4. Lanning, H., "Interface Control Document Communication Performance Margins, No. SE-06", Bendix Systems Division, Ann Arbor, Michigan, 10/4/66.

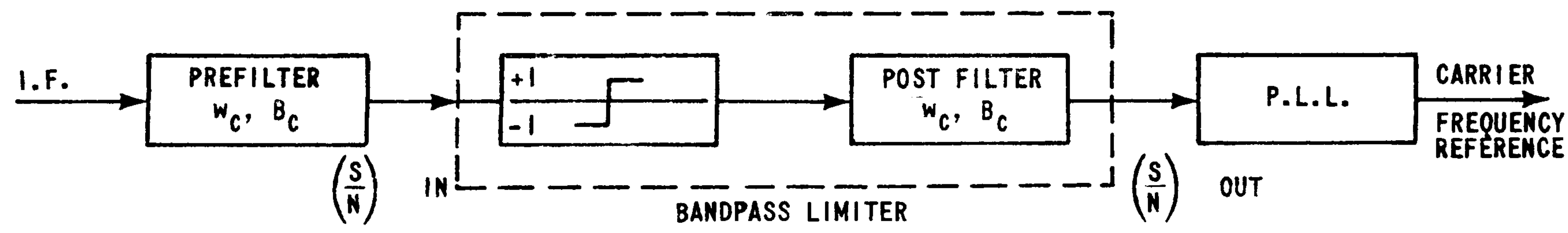

FIGURE 1 - CARRIER CHANNEL

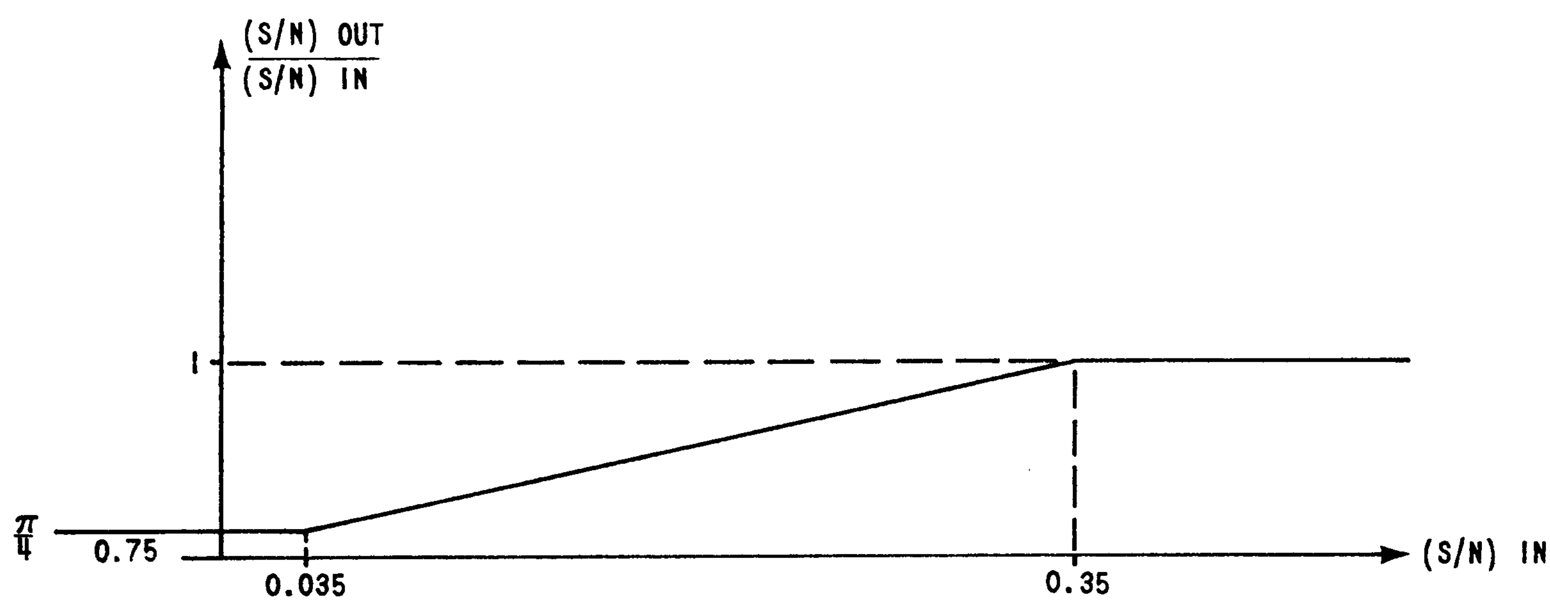

FIGURE 2 - APPROXIMATE $\left(\frac{S}{N}\right)$ TRANSFER CHARACTERISTIC OF BANDPASS LIMITER IN CASCADE WITH A PHASE LOCKED LOOP

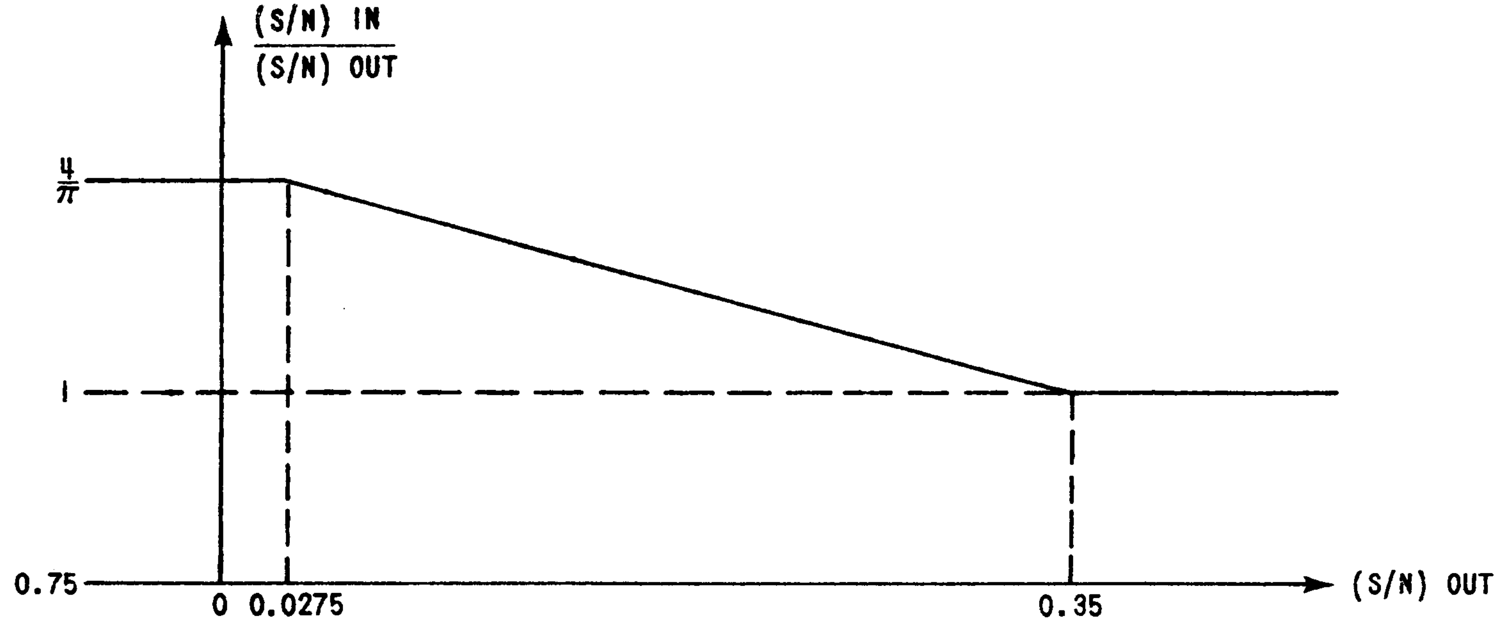

FIGURE 3 - APPROXIMATE $\left(\frac{S}{N}\right)$ TRANSFER CHARACTERISTIC OF BANDPASS LIMITER IN CASCADE WITH A PHASE LOCKED LOOP

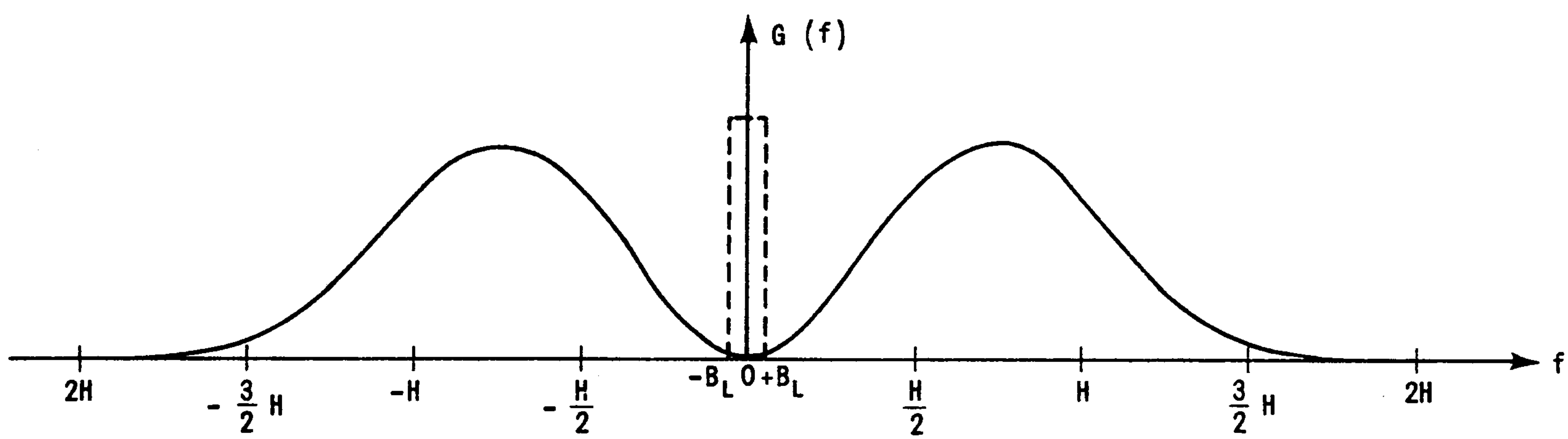

FIGURE 4 - SPECTRUM OF G (f)

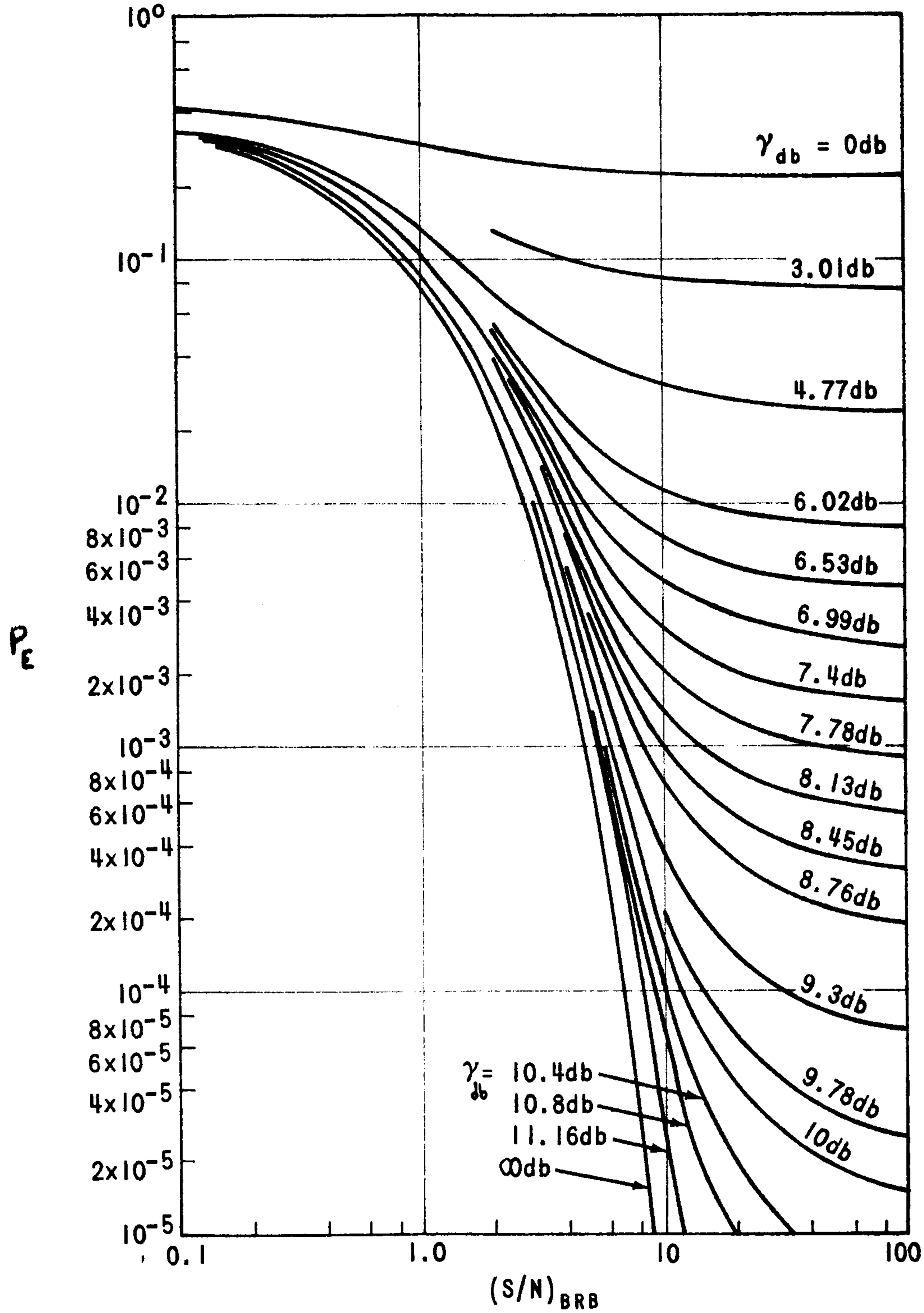

FIGURE 5 - ERROR PROBABILITY P_E VERSUS $(S/N)_{BRB}$ FOR VARIOUS VALUES OF SIGNAL-TO-NOISE RATIO IN SUBCARRIER DATA TRACKING LOOP

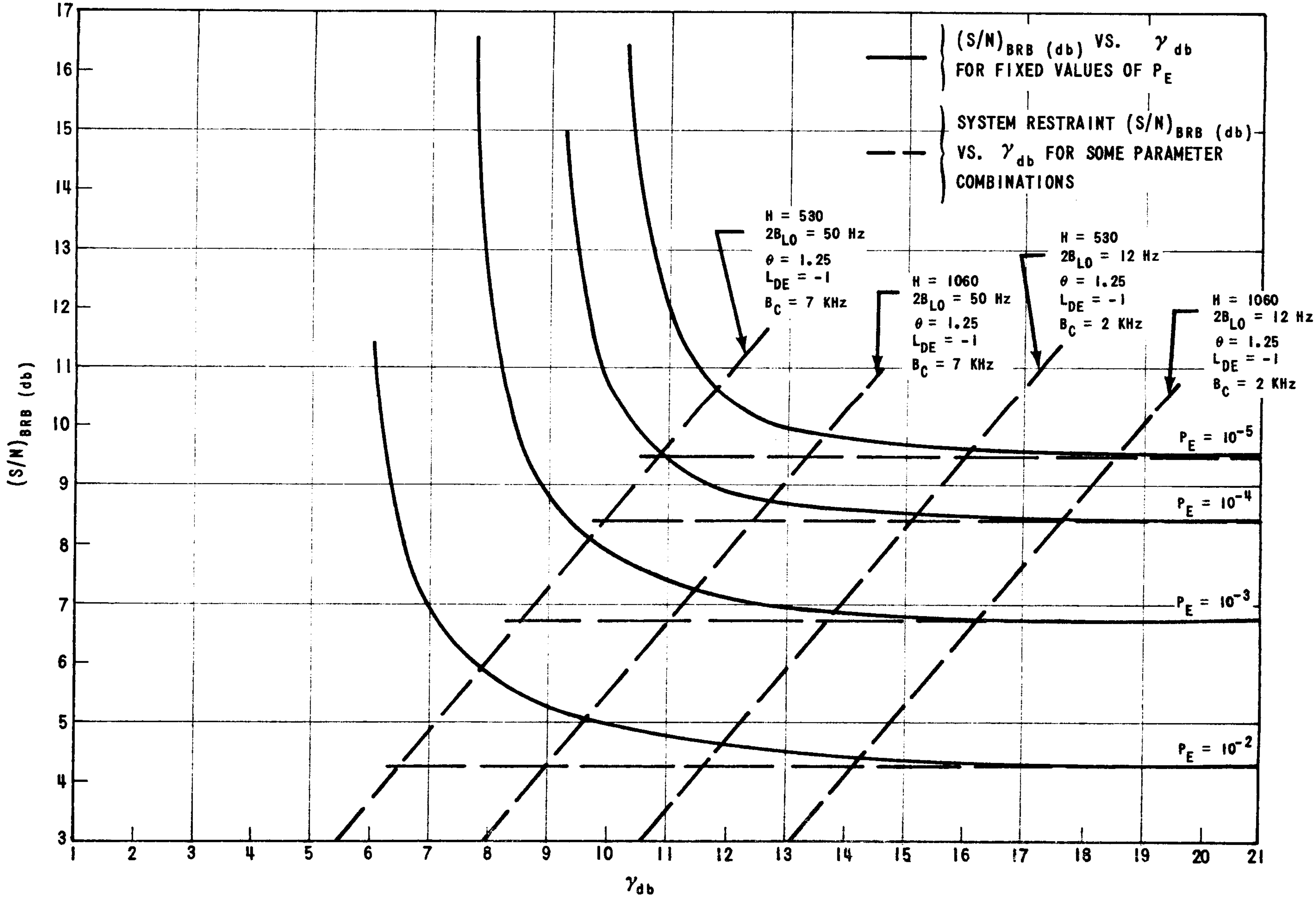

FIGURE 6 - $\left(\frac{S}{N}\right)_{BRB(db)}$ VS. γ_{db} FOR FIXED VALUES OF P_E AND FOR ALSEP - MSFN SYSTEM COMBINATIONS

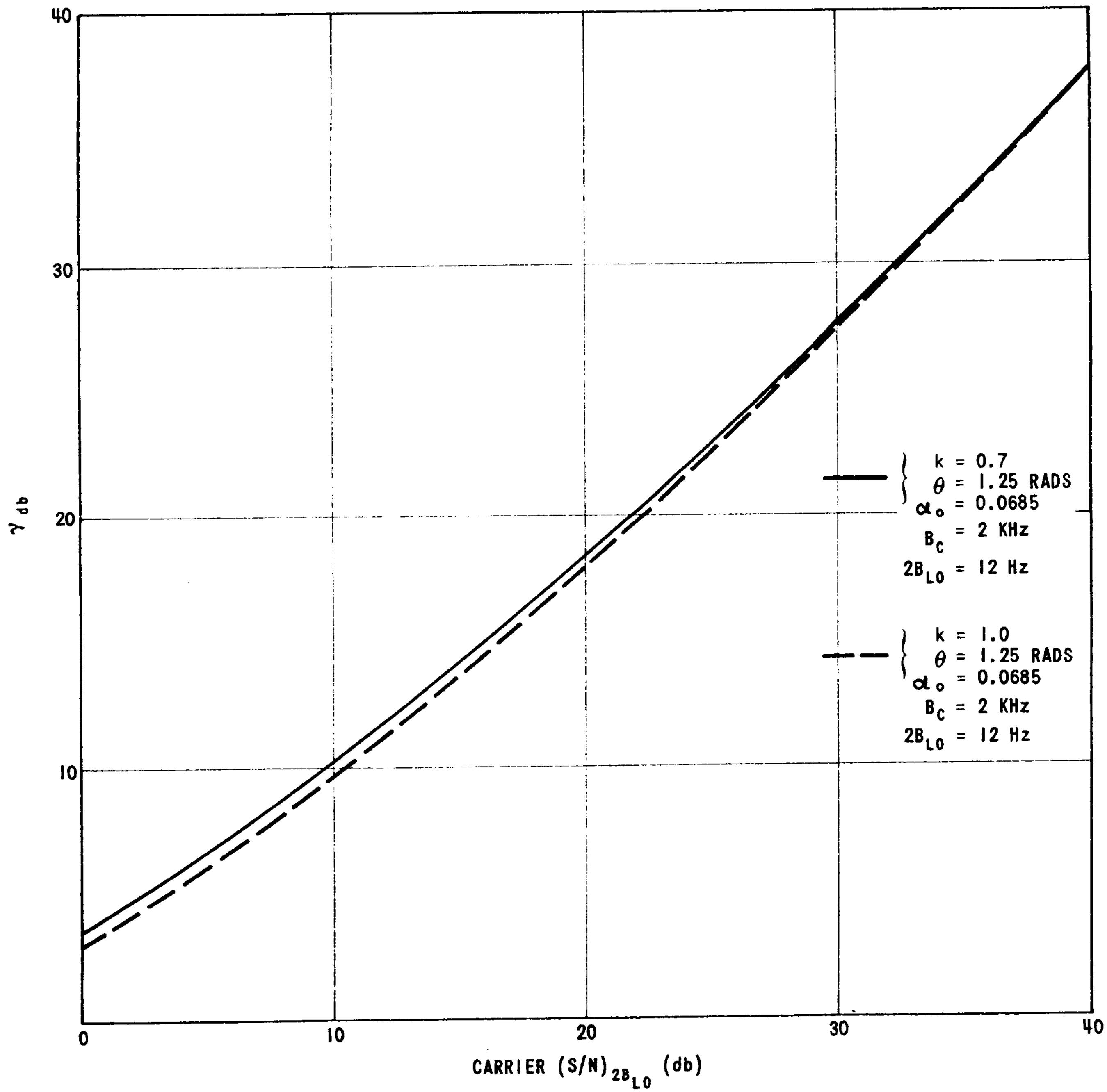

FIGURE 7 - $\gamma_{(db)}$ VS. CARRIER $(S/N)_{2B_{LO}}$

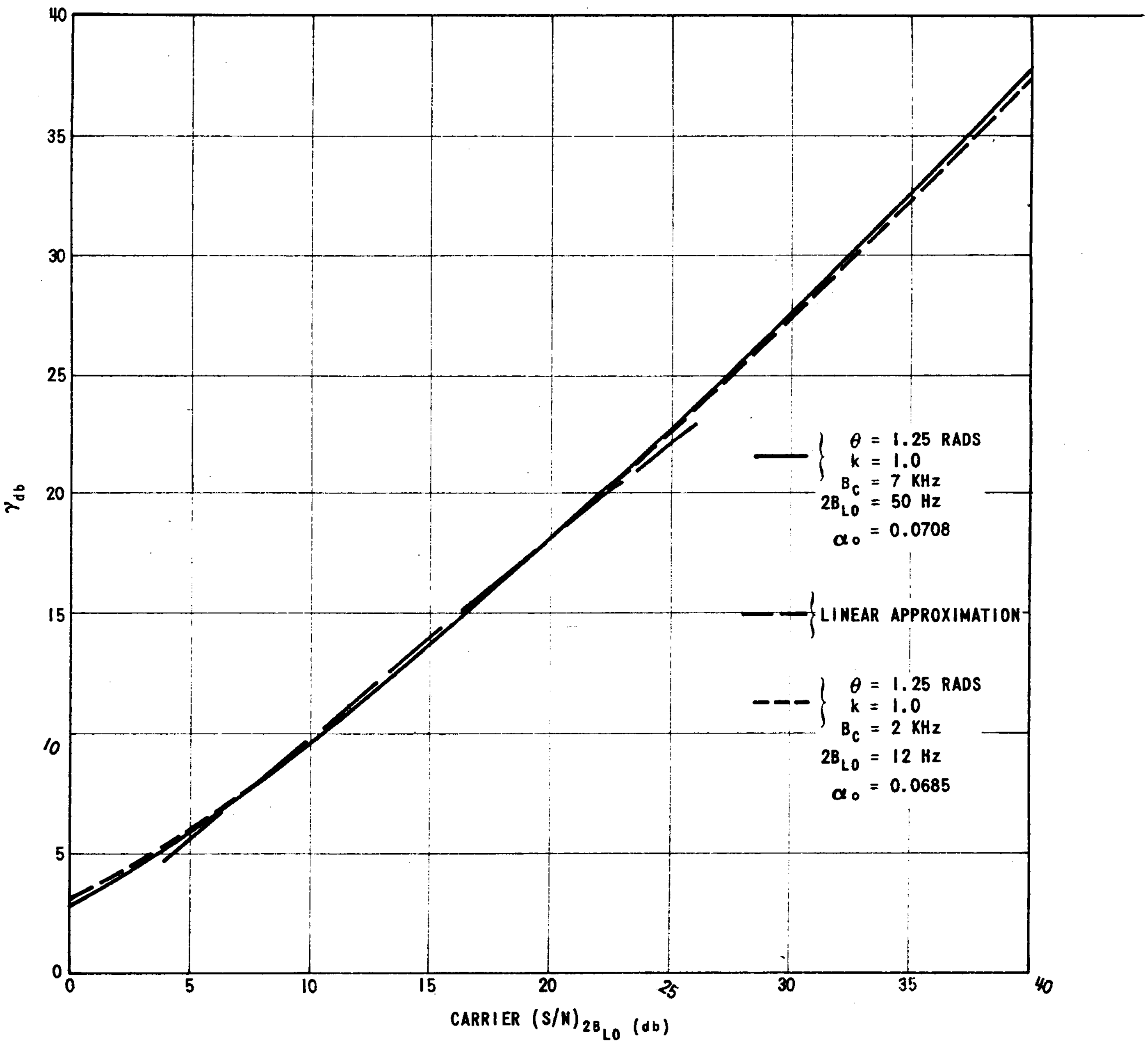

FIGURE 8 - γ_{db} VS. $(S/N)_{2B_{LO}}$ (db)